The Active-Bridge Oscillator*

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Introduction

This paper describes the Active-Bridge Oscillator (ABO), a new concept in high-stability oscillator design. The ABO is a bridge-type oscillator design that is easy to design and overcomes many of the operational and design difficulties associated with “standard” bridge oscillator designs. The ABO will oscillate with a very stable output amplitude over a wide range of operating conditions without the use of an automatic-level-control (ALC). A standard bridge oscillator design requires an ALC to maintain the desired amplitude of oscillation. For this and other reasons, bridge oscillators are not used in mainstream designs. Bridge oscillators are generally relegated to relatively low-volume, high-performance applications. The Colpitts and Pierce designs are the most popular oscillators but are typically less stable than a bridge-type oscillator.

The ABO is a highly integratable circuit and can be designed for use over a very wide frequency range with only minor circuit variations. The high Q of this design can dramatically improve phase-noise, reduce power-supply sensitivity, and minimize circuit related pulling problems common to most popular oscillator designs. The design intention of the ABO is not to make a state-of-the-art bridge-type oscillator but to provide a practical high-performance alternative that delivers most of the performance advantages of a bridge oscillator without the inherent disadvantages of complexity, design difficulty, high-power requirements, and integration potential.

Standard-Bridge Oscillator

Reference material for standard-bridge oscillators (SBO) can be found in papers written by Benjaminson beginning in 1984. Reference 3 contains some historical information on bridge oscillator circuits. For the derivations that follow reference 1 will be used. The derivation of loop-phase is a general case solution where as reference 1 made a simple approximation yielding a specific case solution. Therefore different equations for loop-phase and ultimately loop Q result. Both the special and general case solutions will be described herein.

Figure 1 is a schematic of an SBO. The loop-gain equation for the oscillator is:

\[ A = (\beta_p - \beta_n)A_v = 1 \angle 0^\circ \]  (1)

where A is the loop-gain. For oscillation A must equal 1. Also, \( \beta_p \) must be greater than \( \beta_n \) for the phase condition of oscillation to be met. \( \beta_p \) is defined to be the positive feedback term and \( \beta_n \) is the negative feedback term.

For Figure 1, \( \beta_p = R_2/(R_2 + R_1) \)  (2)

and \( \beta_n = Z_{res}/(Z_{res} + R_f) \)  (3)

\( Z_{res} \) is the impedance of crystal resonator \( Y_1 \). The resonator impedance is be described by the following equation:

\[ Z_{res} = R_m + jL_m\omega + \frac{1}{jC_m\omega} \]  (4)

or \( Z_{res} \) can be written:

\[ Z_{res} = R_m + j\omega^2L_mC_m - \frac{1}{\omega C_m} \]  (5)

where \( R_m, C_m, \) and \( L_m \) are the motional parameters of the quartz resonator. This equation assumes that \( R_m \) is much smaller than the reactive impedance of the resonator shunt capacitance, \( C_0 \). This assumption is very good when using the resonator at series resonance. Since this is a series resonant oscillator, \( Z_{res} \) at series resonance is equal to \( R_m \), the motional resistance of the resonator.

\[ \text{Figure 1. Standard bridge oscillator circuit.} \]
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Hence the negative feedback term at series resonance is:

\[ \beta_n = R_m / (R_m + R_f) \]  

(6)

If there is excess loop-gain and the loop-phase is zero then the nonlinear properties of the gain stage will limit \( A_\ell \) to a value commensurate with equation (1).

The term \( (\beta_p - \beta_n) \) contains all the frequency-dependent loop-phase information. \( \beta_n \) has a phase versus frequency function dependent upon the values of \( R_f \) and \( Z_{res} \), equation (3). This frequency dependent vector is subtracted from a fixed real quantity, \( \beta_p \). The resulting vector, \( (\beta_p - \beta_n) \), determines the oscillator loop-phase. As the magnitude of \( \beta_p \) approaches \( \beta_n \) with \( \beta_p > \beta_n \), \( d\theta / do \) of the resulting vector is greater than \( d\theta / do \) of \( \beta_n \). The penalty paid for this increased phase-rate vector is that the magnitude of this resulting vector decreases, requiring larger \( A_\ell \) to sustain oscillation. Also, as the resulting vector becomes small it is more prone to 1/f noise in the sustaining amplifier.

The circuit loop-phase shift relative to frequency, \( \omega \), is adjustable by choosing the feedback ratios, \( \beta_p \) and \( \beta_n \), such that the desired loop Q is achieved. To determine the effect \( \beta_p \) and \( \beta_n \) have on \( d\theta / do \) of the loop-phase an equation for the loop-phase versus \( \omega \) is required. The loop-phase \( d\theta / do \) will be compared to the to the \( d\theta / do \) of resonator impedance phase to determine the relative loop Q. Let \( M_Q \) be the ratio of the derivatives of oscillator loop-phase to resonator impedance phase with respect to \( \omega \). \( M_Q \) is a measure of the loaded Q of the oscillator in that Q can be equated to the derivative of phase change relative to \( \omega \).

In reference 1 the \( d\theta / do \) for the resonator and the loop-gain are derived for figure 1. Resonator Q was correctly equated to \( d\theta / do \) in reference 1 by deriving the \( d\theta / do \) of the resonator impedance \( Z_{res} \) phase to determine the relative loop Q. Let \( M_Q \) be the ratio of the derivatives of oscillator loop-phase to resonator impedance phase with respect to \( \omega \). \( M_Q \) is a measure of the loaded Q of the oscillator in that Q can be equated to the derivative of phase change relative to \( \omega \).

The following equation is the derivative of the resonator impedance phase with respect to \( \omega \):

\[ \frac{d\theta_{Resonator}}{d\omega} = 2Q \frac{2L_m}{\omega_s} \frac{1}{R_m} \]  

(7)

where \( \omega_s \) is the series resonance of the resonator.

\[ \omega_s = \frac{1}{\sqrt{L_mC_m}} \]  

(8)

and \( f_s = \omega_s / 2\pi \).

The loop-gain \( d\theta / do \) is found by taking the derivative of the phase of \( (\beta_p - \beta_n) \) with respect to \( \omega \). The loop-phase equation is determined first by finding the phase of the following:

\[ (\beta_p - \beta_n) = \beta_p - \frac{Z_{res}}{Z_{res} + R_f} \]  

(9)

Substituting the motional arm impedances for \( Z_{res} \), \( (\beta_p - \beta_n) \) becomes:

\[ (\beta_p - \beta_n) = \beta_p - \frac{R_m + j \frac{\omega^2 L_m}{\omega C_m} - 1}{R_m + j \frac{\omega^2 L_m}{\omega C_m} + 1 + R_f} \]  

(10)

Let \( \theta_{loop} \) be the phase of the expression \( (\beta_p - \beta_n) \). After a little algebra, \( \theta_{loop} \) is:

\[ \theta_{loop} = \tan^{-1} \left( \frac{\frac{\omega^2 L_m}{C_m} - 1}{\beta_p (R_m + R_f) - R_m} \right) - \tan^{-1} \left( \frac{\omega^2 L_m}{\omega C_m} \right) \]  

(11)

The derivative of \( \theta_{loop} \) with respect to \( \omega \) was found with a symbolic math program. The solution of \( d\theta_{loop} / d\omega \) is lengthy and is simplified greatly by using the assumption that the oscillator operates at \( \omega_s \). At \( \omega_s \), \( (\omega^2 L_m C_m - 1) / \omega C_m = 0 \) and a relatively simple equation results. The following is the expression for \( d\theta_{loop} / d\omega \) at \( \omega_s \):

\[ \frac{d\theta_{loop}}{d\omega} = -2L_m \frac{R_f}{(R_m + R_f) \beta_p (R_m + R_f) - R_m} \]  

(12)

The goal of this exercise is to find the expression for \( M_Q \), which is the ratio of \( d\theta_{loop} / d\omega \) to \( d\theta_{Resonator} / d\omega \). The following is the derived expression for \( M_Q \):

\[ M_Q = \frac{\beta_n - 1}{\beta_p - \beta_n} \]  

(13)

Reference 1 derives a different but similar equation for \( M_Q \), equation (14). First, the sign of the two equations is opposite because reference 1 derived the phase of \( 1/(\beta_p - \beta_n) \). Upon careful examination of this derivation I also conclude that \( \beta_n \) was made proportional to \( R_m \) for reasons of simplification. If \( \beta_n \) is made proportional to \( R_m \) then this implies that \( R_f >> R_m \). This assumption simply forces \( \beta_n << 1 \).

\[ M_Q = \frac{\beta_n}{\beta_p - \beta_n} \]  

(14)

This \( M_Q \) is not the same as equation (13). When one assumes \( \beta_n << 1 \) then the magnitude of equation (13) does approach the magnitude equation (14), which would be consistent with my observations and conclusions. These facts explain and help verify the claim that equation (13) is a general case solution for the bridge.
oscillator described in figure 1. Also, these results have been verified with SPICE analysis and negative resistance modeling. For any further reference to $M_Q$ equation (13) is the assumed definition.

It is interesting to note that if $\beta_p$ is equal to one, than $M_Q = \beta_n$. If $M_Q = \beta_n$ then the loop $Q$, $d\phi/d\omega$, can never be made higher than the $Q$ of the resonator. This is because $\beta_n$ can only approach one when $R_t$ approaches zero. Although mathematically possible, this situation is impractical to say the least. By making $\beta_p$ a little larger than $\beta_n$, the circuit $M_Q$ can be made $\geq 1$. By proper selection of $\beta_p$ and $\beta_n$, $M_Q$ may be designed for most any desired value around one with ease. Figure 2 graphically compares the loop-phase of a hypothetical 438 kHz oscillator with the resonator phase as a function of frequency. The feedback terms were: $\beta_n = 0.2$ and $\beta_p = 0.2857$. The resulting amplifier gain and $M_Q$ were calculated to be: $A_v = 11.6$ and $M_Q = 1.87$. In this graph the phase of the resonator was inverted to better see the relative slopes of both curves, $d\phi/d\omega$. Note that at the oscillation frequency the slope of the oscillator loop-phase is greater than that of the resonator impedance phase by a factor of 1.87.

These two graphs demonstrate that the phase-slope of the oscillator loop is larger than the resonator phase slope and that the loop gain is at a local minimum at the oscillation frequency. Typically, non-bridge oscillators operate with the loop-gain at a local maximum.

![Figure 2. Oscillator loop-phase and resonator impedance phase versus frequency for $M_Q=1.87$.](image)

![Figure 3. Oscillator loop gain versus frequency, $M_Q=1.87$.](image)

**Practical Realization of the SBO**

Figure 4 is a practical realization of an SBO. This design uses $Q_1$ and $Q_2$ as the differential amplifier gain, $A_v$, where the collector of $Q_1$ drives the bridge-network. The bridge-network is comprised of $C_1$, $C_2$, $R_6$ and $Y_1$.

![Figure 4. Schematic of a standard bridge oscillator.](image)

$L_1$ is an RF choke that provides dc bias to the base of $Q_1$ equal to the base of $Q_2$. $C_3$ is a large value dc blocking capacitor. The tank in the collector of $Q_1$ ($L_c$, $C_0$) is used to make the collector impedance real at the desired resonator frequency. This SBO has some
inherent shortcomings and does not well represent the ideal bridge oscillator depicted in the “block” schematic, figure 1. First, this design uses a high-impedance collector node (Q1) to drive the bridge network. This network uses C1 and C2 to provide βp. Rf and Y1 provide βn. The differential amplifier gain is a function of the collector impedance and the transistor h where:

\[ h = \frac{kT/q}{I_e} = 1/g_m \]  
(15)

at 25°C \[ h = \frac{26mV}{I_e} = 1/g_m \]  
(16)

and Ie is the dc current in the transistor. The gain of the differential amplifier, Av, can be expressed as:

\[ A_v = \frac{Z_{\text{collector}}}{2h} \]  
(17)

where \( Z_{\text{collector}} \) is the impedance from collector to ground of the Q1 transistor. Since Rf and the resonator directly load the collector the following equations can be used to describe the feedback and gain characteristics of the SBO in figure 4:

\[ A_v \approx \frac{R_f + R_m}{2h} \]  
(18)

\[ \beta_n = \frac{R_m}{R_m + R_f} \]  
(19)

\[ \beta_p = \frac{1}{1 + \frac{C_2}{C_1}} \]  
(20)

The Av term is only an approximation and assumes no loading by the tank circuit (Cno, Lno, C1 and C2) or the bias resistors R1 and R2. Note that the tank circuit includes the βp structure, C1 and C2. βp is provided by the capacitive voltage divider between C1 and C2, equation (20). This equation assumes that R1 and R2 are much greater than the reactive impedances of C1 and C2.

To obtain relatively high M0 it is desirable to make Rf relatively small, which provides significant negative feedback, βn equation (19). If Rf is made small Av is decreased, from equation (18). The only way to overcome this loss in gain is to significantly increase the bias current in the differential amplifier transistors to decrease h, therefore increasing Av. For ease of design and relative current efficiency it may be preferable to not have Av a function of the feedback parameter impedances, in particular Rf. The only way this is achievable is to have a low impedance node at the collector Q1.

Another disadvantage with the oscillator in figure 4 is that only the differential amplifier nonlinearities, h, control the amplitude of oscillation. As a transistor base-emitter ac voltage increases so does the effective nonlinear value of h. For high M0 operation, the signals at each base are approximately the same amplitude. This means that the ac base-to-emitter voltages of the differential amplifier transistors are relatively small even for large signals at each base. Therefore, large changes in amplitude result in only small changes in operating h. This means that the oscillation amplitude will vary greatly with respect to resonator resistance, temperature, etc. Therefore this circuit must be used with an automatic level control (ALC) circuit which controls some element in the oscillator (bias or feedback component) continuously to maintain a desired signal amplitude.

The Active-Bridge Oscillator

The active-bridge oscillator (ABO) was discovered while modeling a half-bridge design. A half-bridge design is shown in figure 5.
decreased the half-bridge design turns into a full-bridge function where $\beta_n$ is now less than one. $\beta_p$ becomes a function of this emitter to ground impedance and the transistor $h$ in this non-obvious bridge circuit.

Figure 6 is the simplified active-bridge oscillator schematic used as a derivation reference.

![Figure 6. A simplified ABO schematic.](image)

The loop equations can be derived by breaking the oscillator loop at the emitter of $Q_3$, base of $Q_2$ node and determining the loop-gain. This is a logical place to break the node and preserve the oscillator closed-loop characteristics. This is because the emitter of $Q_3$ is low impedance, and the $\beta_p$ and $\beta_n$ functions remain intact.

This circuit is designed with $Q_3$ used as an emitter follower to buffer the collector $Q_1$ from the low impedance $\beta_n$, and serve as an output buffer. $\beta_n$ comprises $R_e$ and $Y_1$. $Y_1$ is equal to $R_m$ at $\omega_0$. $R_f$ also supplies dc bias to the bases of $Q_1$ and $Q_2$. This is possible because $R_f$ is typically a relatively low value resistor, at most hundreds of ohms. $R_e$ is low value resistor, tens of ohms (ac). $R_e$ used in conjunction with the intrinsic emitter impedances of the transistors, $h$, to provide the $\beta_p$ function. Hence the name active-bridge oscillator.

In deriving the loop gain, it was found to fit the form of a bridge oscillator function, equation (1), with the following characteristics. For simplicity these equations assume relatively large transistor $b$.

$$A_v = \frac{R_c}{h + \frac{hR_e}{R_e} + h}$$

$$\beta_n = \frac{R_m}{(R_m + R_f)}$$

$$\beta_p = \frac{R_e}{(R_e + h)}$$

$\beta_p$ and $A_v$ are functions of $h$ and $R_e$. It is interesting to note that as $R_e$ is decreased, increasing $M_Q$, that $A_v$ is increased. Since increasing $M_Q$ requires a larger $A_v$ to sustain oscillation this characteristic is desirable from a current efficiency standpoint. Essentially, you get a little 'free gain' when it is required.

Figure 7 is a practical example of the ABO with consideration given to bias and filtering necessities.

![Figure 7. The complete active-bridge oscillator design.](image)

This circuit includes $R_1$ and $C_2$ to facilitate dc biasing and provide an ac ground for $R_e$. Capacitors, $C_3$ and $C_4$, are dc blocking capacitors and considered as ac shorts. $R_e$ provides $\beta_p$ and is therefore a low value (typically $< 100\Omega$) where $R_1$ is a bias resistor and typically set from $1 \, k\Omega$ to $2 \, k\Omega$. The bias voltages in the circuit are therefore primarily determined by $V_{cc}$, transistor $V_{be}$, $R_1$ and $R_e$. Equation (24) is the dc bias equation for the dc voltage $V_e$ (figure 7) relative to ground. This equation is valid only when the circuit is not oscillating. When the circuit oscillates there is significant bias shifting in the circuit. $V_e$ is the dc voltage at the collector of $Q_1$ when $R_e$ is much less than $R_1$.

$$V_{cc} + 2V_{be} = \frac{R_e}{2R_1}$$

$$V_e = \frac{1 + \frac{R_e}{2R_1}}{2R_1}$$

This equation can be greatly simplified if $R_e = R_1$, which is a convenient and practical design possibility for this circuit.
\[ V_c = \frac{V_{ce} + V_{be}}{1.5} \]  
when \( V_c \) is set equal to \( R_1 \), and \( R_s < R_1 \).

The emitters of the differential amplifier \((Q_1, Q_2)\) will be biased at two \( V_{be} \) drops down from \( V_c \) due to the base-emitter drop of \( Q_3 \) and \( Q_2 \). \( R_2 \) sets the bias current in \( Q \). The dc voltage at \( V_{be} \) is simply \( V_c \) minus \( V_{be} \). Therefore the bias current for \( Q \) is \( \frac{V_{be}}{R_2} \). This bias scheme works nicely in that it is very simple yet biases all transistors simultaneously at reasonable voltages and currents. The tank circuit, \( L_C \), performs two functions. First, for efficient use of current and a high available \( V_o \), the collector resistor \( R_c \) is chosen to be relatively large. Any significant circuit capacitance at this node will make the phase of the collector impedance a negative phase angle. For proper circuit operation all feedback terms and \( A_v \) need to be real. The tank, \( L_C \), allows \( A_n \) to be made real at any desired frequency. Secondly, this tank significantly reduces the distortion in the output waveform by shunting the undesired harmonics of the oscillator to ground. Also, this tank naturally selects the desired resonator overtone. Overtone operation at \( f_e \) is as easily achieved as fundamental mode operation.

Due to the predominately resistor-transistor based construction of this design, an analog ASIC would allow for a very small oscillator circuit that provides ease of design and high-performance over a wide range of operating frequencies. Reasonable values for an ABO when \( R_c = 20 \Omega \) are: \( R_e = R_1 = 1.5 \Omega \), \( R_p = 75 \Omega \), \( R_t = 100 \Omega \) and \( R_2 = 400 \Omega \). \( L_e \) and \( C_e \) are chosen to be parallel resonant and approximately 100 \( \Omega \) to 150 \( \Omega \) of reactive impedance. This results in an oscillator with high operating \( Q \), low-distortion and a low-output drive at \( Q_3 \). This oscillator designed for 20 MHz operation delivers approximately minus 2dBm into a 50 \( \Omega \) load. The waveform is a very low distortion (2\(^{nd}\) harmonic -30 dBc) sine wave. This circuit required approximately 10 mA of current when designed for a \( V_{be} \) of 5 V. \( Q_3 \) is the major power consuming component in that it must drive the low impedance \( R_t \) and \( R_{load} \). To do this the bias in \( Q_3 \) is relatively high.

**Active-Bridge Nonlinearities**

The ABO steady-state, large-signal oscillation amplitude can be made relatively independent of resonator resistance, \( V_{be} \), and temperature changes. This is due primarily to two factors. First, the steady-state base signals on the ABO are not very similar in amplitude as they are in a SBO. The base of \( Q_3 \) is the same amplitude as the collector signal at \( Q_1 \). The base of \( Q_1 \) is being driven by a fraction of the collector of \( Q_1 \). This allows the individual transistor \( h \) to be a strong function of signal amplitude. As the oscillation amplitude increases \( h \) increases. Secondly, the \( \beta_p \) term of the ABO is a strong function of oscillation amplitude, \( \beta_p = R_c/(R_e + h) \) where \( h \) is the nonlinear transistor-impedance term. As a result the oscillation amplitude is relatively stabilized by the circuit \( M_Q \) via transistor \( h \). Therefore slight changes in the circuit parameters, such as resonator resistance, do not effect the oscillation amplitude greatly.

These characteristics are useful in that the ABO does not require an ALC for low distortion large-signal sinusoidal outputs.

Figure 9 is a graph of a calculated \( M_Q \) versus transistor \( h \) for the following hypothetical ABO design: \( R_c = 1100 \Omega \), \( R_p = 105 \Omega \), \( R_m = 20 \Omega \), \( R_e = 75 \Omega \). Thus \( \beta_n \) calculates to 0.160. Using equation (1) and the oscillator feedback and \( A_v \) terms, equations (21), (22), and (23) the steady-state value for \( h \) was determined. The calculated value of \( h \) is 143 \( \Omega \). Therefore \( A_v = 5.6 \), \( \beta_p = 0.338 \) and \( M_Q \) calculates to 0.753.

![Figure 9. \( M_Q \) versus \( h \) for \( h \) equals 15 to 250 ohms.](image)

Figure 10 is a graph of \( A_v \) versus \( h \) for the given hypothetical ABO. This graph demonstrates that \( A_v \) is decreasing with increasing \( h \). It can also be seen in this graph that \( A_v \) is 5.6 at \( h = 143 \Omega \). This value of \( A_v \) is the steady state gain. \( A_v \) is equal to \( 1/(\beta_p \beta_n) \), as equation (1) states.
Figure 10. $A_v$ versus $h$, for $h$ equals 15 to 250 ohms.

To demonstrate the interdependence of $A_v$ and the feedback terms this hypothetical oscillator circuit will remain constant and the resonator resistance will be allowed to vary slightly. Let the resonator resistance, $R_m$ increases from 20 $\Omega$ from 25 $\Omega$. The oscillation amplitude will decrease just enough to obtain a new steady-state value of $h$ that satisfies the loop-gain equation (1). In this case $h = 134 \Omega$ and the resulting $M_Q$, feedback terms, and $A_v$ are: $M_Q = 0.936$, $\beta_p=0.358$, $\beta_n=0.192$ and $A_v=6.02$. Note that $\beta_p$ changes from 0.338 to 0.358 for $R_m$ going from 20 $\Omega$ to 25 $\Omega$. $M_Q$ changes by 24.3% for this 5 $\Omega$ change in $R_m$. This relatively small change in $M_Q$ is due to $\beta_p$ increasing with $\beta_n$ via the change in transistor $h$.

It is interesting to compare a hypothetical SBO (like Figure 4) to the ABO with the same feedback parameters allowing the resonator $R_m$ to vary from 20 $\Omega$ to 25 $\Omega$. Let $\beta_p=0.358$, and $R_f=105 \Omega$ and $R_c=1100 \Omega$ but assume that the collector impedance ($Q_1$) is buffered by a transistor (as in the ABO) such that the $A_v$ for the hypothetical SBO is $R_c/2h$. This allows a fair comparison of the two designs in that both circuits now buffer the bridge-network from the high impedance collector node.

When this sensitivity study is evaluated one finds that for an $R_m$ change of 20 $\Omega$ to 25 $\Omega$ that $M_Q$ goes from 0.7534 to 1.0632, a 41% change. Transistor $h$ goes from 98.12 $\Omega$ to 80.35 $\Omega$. The percentage change in $h$ required by the SBO to absorb this change in $R_m$ is approximately six times greater than for the ABO. $M_Q$ for the ABO is more stable over this change in $R_m$ than an SBO. Also, since the “amplitude versus h” function is much steeper with the ABO than an SBO (as previously described) the ABO amplitude stability is much greater than that of an SBO.

When these two oscillators were built and evaluated for $R_m$ varying from 20 $\Omega$ to 25 $\Omega$, the SBO oscillation-amplitude changed 33%, and the ABO amplitude changed only 5 %.

**Active-Bridge Experimental Data**

The ABO of Figure 11 was used to verify the ABO loop equations for $M_Q$. For this experiment $R_c=1100 \Omega$, $R_f=1500 \Omega$, $R_m=20 \Omega$, $R_{load}=200 \Omega$. $R_c$ was actually 1500 $\Omega$ but due to tank-loading and the finite impedance of the buffer, $R_c=1100 \Omega$ was used for the
calculations. A 20 MHz resonator was used for this test. The circuit M_Q was varied by changing R_e from 35 Ω to 75 Ω. The M_Q was determined by measuring the frequency shift of the oscillator when a small capacitor, 1pF, shunted the collector load of Q_1. This added capacitance places a small, known, phase-shift into the loop. Measuring the frequency shift required to accommodate the collector phase-shift measures the resonator phase-shift. These two phase-shifts allow for a determination of M_Q. The collector phase-shift for a 1 pF delta capacitance across R_e is approximately 7°. The resonator phase-slope measured 114.3 m°/Hz, near series resonance.

Therefore: \[ M_Q_{\text{measured}} = \frac{\Delta \text{FreqOsc}}{114.3 \text{m°} / \text{Hz}} \] (26)

where ΔFreqOsc is the frequency shift of the oscillator due to a 1pF capacitance placed in parallel with the collector tank of Q_1. Table 1 displays the results of the described experiment.

<table>
<thead>
<tr>
<th>R_e</th>
<th>M_Q measured</th>
<th>M_Q calculated</th>
</tr>
</thead>
<tbody>
<tr>
<td>35 Ω</td>
<td>1.33</td>
<td>1.23</td>
</tr>
<tr>
<td>51 Ω</td>
<td>1.13</td>
<td>0.957</td>
</tr>
<tr>
<td>75 Ω</td>
<td>0.87</td>
<td>0.75</td>
</tr>
</tbody>
</table>

Table 1. Data table for the M_Q experiment of an ABO.

These data demonstrate that the loop-phase indeed can be made steeper than the resonator slope, when R_e equals 35Ω. This data also demonstrates how well the large signal operation of the design can be predicted from the derived equations which are large signal approximations using small signal parameters, h. The oscillation amplitude decreased approximately 40% for R_e going from 75 Ω to 35 Ω. This is very good performance considering the M_Q varies approximately 53% for this range of R_e.

The ABO With Fixed Feedback Terms

There are applications for bridge-type oscillators where it is undesirable to have d0/df a function of transistor h. Figure 12 shows a schematic of an ABO that is being used for a quartz-sensor application where the ABO function has a fixed value for β_p.

The impedance of Q_7 is substituted for the R_e term in the ABO loop-gain terms, A_V and β_p, of equation (21) and (23). Transistor Q_7 is configured as a diode. This diode impedance value is equal to the intrinsic emitter impedance. Q_1 and Q_2 are defined here to have intrinsic emitter impedance equal to h. Q_7 is biased at of twice that of Q_1 (or Q_2) making the impedance of Q_7 equal to h/2. Therefore, this oscillator has the following loop-gain terms where R_e is assumed to dominate the collector to ground impedance:

\[ A_V = \frac{R_e}{1.5h} \] (27)

\[ \beta_p = \frac{R_m}{(R_m + R_f)} \] (28)

Since \[ \beta_p = \frac{(h/2)/(h/2)+ h} \]

\[ \beta_p = 1/3 \] (29)

β_p is now a constant and β_n is as before, a voltage divider between R_f and R_m. This circuit biases the collector of Q_1 to 3 V_b drops above ground. Q_6 sets the current in the differential pair, Q_1-Q_2, by current mirroring. The bias current in Q_6, controlled by Q_5 and R_1 is mirrored in Q_4. Due to the low value R_e, the bias current of Q_1 and Q_2 are equal to the current in Q_4.

Figure 12. The ABO with fixed value feedback terms.
This current directly controls the magnitude of \( A_c \), via \( h \), equation (27). Thus an ALC circuit can control the amplitude of oscillation of node \( V_{out} \) via a control voltage, \( V' \). This design is currently being used to measure very small changes in resonator resistance where a quartz resonator is being used as a sensing element. In this application both frequency and \( R_m \) information are required. It is desired to resolve very small changes in \( R_m \). This circuit has resolved milli-ohm changes in a nominal 25 \( \Omega \) resonator. The original ABO circuit tends to mask these changes because \( \beta_r \), is not a constant, as described earlier in this paper. Also to maintain circuit linearity, this design is being used at very small oscillation amplitudes. In this situation, the design behaves very similar to an ideal SBO. The main advantage of this design over an SBO is the circuit integration potential and the simplicity of the circuitry.

**Conclusions**

General bridge-oscillator design equations were derived and discussed as an introduction to a new oscillator, the Active-Bridge oscillator. The ABO feedback and gain terms were presented and used in hypothetical and actual design examples. These examples included comparisons to previous work, references 1 and 2.

This design is both practical, and in many ways an excellent general use oscillator that is good for high-stability applications at most any frequency that quartz resonators are constructed. ABO designs have been implemented easily with 125 MHz fifth overtone resonators or with fundamentals from 5 MHz to 200 MHz with essentially the same design! With high \( f_0 \) transistors or ASIC implementations, higher frequencies could easily be achieved.

Another significant advantage of the ABO is that it operates with the resonator at series resonance. This is a distinct advantage at high-frequencies where manufacturing resonators tuned to \( f_0 \) is easier and more consistent than when tuned to a given load capacitance. This is true because the ill effects of the resonator \( C_0 \) (shunt capacitance) are minimized with \( f_0 \) tuning. At \( f_0 \) the resonator is near a local low-impedance point which is relatively easy to measure and not very dependent on the value of \( C_0 \).

The ABO also produces a low-distortion sine wave output into low impedances with no additional filtering or buffering. Very efficient low-current designs are possible with the ABO using the class-AB follower. With \( V_{cc} \) as low as 3 Volts designs drawing 2mA at 20 MHz were easily achieved. The bias arrangement and nonlinear properties of the ABO allow operation over a wide temperature range with little or no amplitude variations. The design uses a grounded resonator which greatly simplifies the design of VCOs (voltage controlled oscillator) or TCXOs (temperature compensated crystal oscillator) that require some form of “pulling” circuitry such as varactor networks. The phase-noise of the design has been found to be at least as good as a well designed Pierce oscillator using the same resonator. This design is still in development but it appears that the close-in noise is excellent and the far out (>10 kHz) noise is about 10 dB worse than a typical single transistor Pierce or Colpitts design. This is most probably due to the relatively large number of active components used to implement the oscillator. But, oscillator pulling due to temperature, power supply, or load, are superior to any oscillator (without an ALC) the author has yet encountered. The overall performance and design flexibility make this design a relative breakthrough in that it is nearly a “universal” design and can be used for many applications and frequencies.

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**References**


